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THEORETICAL DISCUSSION OF DIPHASE TRANSMISSION

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September 1963

UNITED STATES ARMY
ELECTRONICS RESEARCH AND DEVELOPMENT LABORATORY
FORT MONMOUTH, N.J.

U. S. ARMY ELECTRONICS RESEARCH AND DEVELOPMENT LABORATORY

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THEORETICAL DISCUSSION OF DIPHASE TRANSMISSION

L. U. Dworkin

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Abstract

A differential diphase repeater, initially developed by O. Ringlehaan, was subjected to extensive theoretical analysis. In this report dipulse and diphase systems are compared with respect to their power density spectrum, detectability, and tolerance to noise and timing extraction. Results verify that the diphase system has a higher noise threshold, less timing jitter and is a better all-around method of transmission. In addition, tests of the initial repeater models indicate that they are simpler and have lower power consumption than the present dipulse repeaters.

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THEORETICAL DISCUSSION OF DIPHASE TRANSMISSION

INTRODUCTION

The problem of sending many communication channels over a single cable for relatively great distances has long been one of the primary problems in military communications. This report deals with a theoretical evaluation of a digital modulation approach that can be used to greatly improve PCM transmission over Cable, CX-4245. The present cable transmission system consists of attended repeater points every 40 miles with unattended repeater points every mile. These repeaters must carry out total pulse regeneration for signals at a pulse rate of 2.304 megabits per second, corresponding to 48 digitized voice channels. The present modulation approach, which is used to send PCM over a cable, is called dipulse and has a variety of deficiencies associated with its realized system. Some of these include accumulation of timing jitter, relative sensitivity to noise, and expense of the unattended repeaters. What we are proposing is the use of diphase modulation in place of dipulse with no change in repeater spacing or pulse transmission frequency. The diphase system was investigated because it offered a method of reducing unattended repeater cost, increased noise threshold and also reduced timing jitter. In the context of this report a detailed discussion of the cable system, the problem associated with digital transmission, and comparison of various modulation approaches will be considered. Most of the discussion will deal with differential diphase modulation and its characteristics.

BACKGROUND

The material presented in this section is a collection of information associated with digital repeater design. Some of the common problems in connection with binary transmission and a variety of different repeater approaches are briefly discussed.

1. Repeater Description

a. The regenerative repeater that will be discussed has the function of totally reconstructing a digital signal. This involves reproducing an exact replica of the original transmitted pulses after each repeatered link. In most applications, the cost imposes the requirement that repeater spacing be of a magnitude so as to cause the transmitted train to be severely distorted by the cable transfer characteristic. Each repeater must amplify, reshape, and retime the distorted pulse train. A typical block diagram of a regenerative repeater is shown in Fig. I.

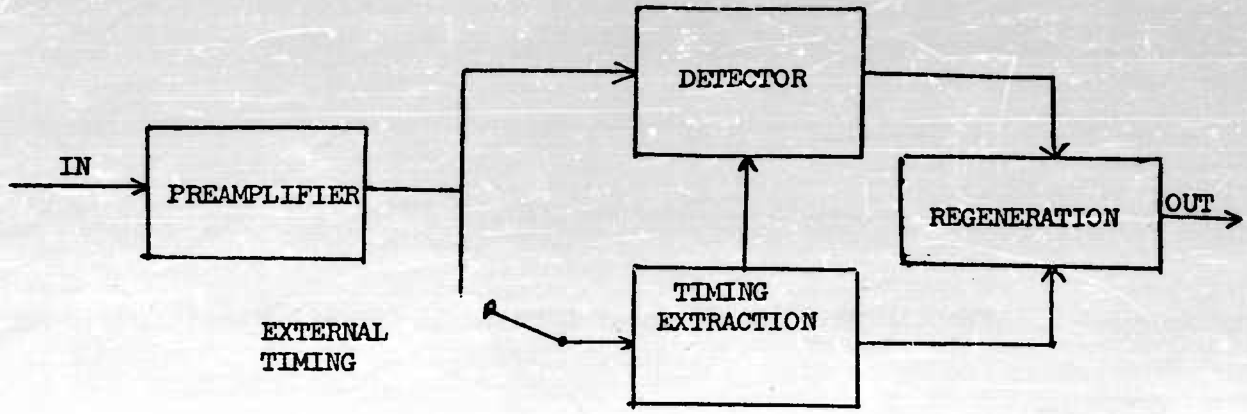


FIGURE I. Digital Repeater

b. The preamplifier block carries out initial amplification as well as equalization. The revitalized pulses are then identified as a "1" or a "0," using threshold detection. In order to detect successfully, timing information is required and this can be obtained from the pulse train itself or some external source. Finally, the detector decision and timing are used to regenerate the pulses so their shape is the same as it was before transmission. If this system were ideally designed, there would be no limit to the length of an overall system. However, practically speaking, this is not the case. A number of problems occur with primary difficulty in timing recovery. For this reason we will devote a major portion of our work to this topic.

2. <u>Digital Reconstruction</u>

a. <u>Types of Timing</u>

(1) When considering the task of retiming, two primary operations are involved:

(a) Timing extraction

(b) Timing utilization

(2) Timing extraction refers to the recovery of the timing signal from the transmitted train while timing utilization involves employing the extracted signal in the detection and regeneration processes. First, we will consider timing extraction.

(3) If auxiliary timing is made available, the received pulse train is not used for timing extraction. This approach is called external timing. Here, we will only consider self-timed approaches where the timing is extracted from the received pulse train itself. In addition, we will also eliminate the possibility of adding a periodic component to the binary pulse train. Our consideration is limited to the binary pulse train and whatever timing information may be associated with the train itself. Where this is the case, there are two primary repeater configurations used to achieve a timing signal. The first of these, called forward retiming, allows the extraction

of the timing signal from the revitalized pulses out of the preamplifier (FIG. I). The second, known as backward retiming, uses the regenerated pulse train at the output of the repeater for timing extraction and feeds back this information for use in regeneration. A full discussion of both of these approaches can be found in Ref. No. 1, 2 and 3. For the most part, forward retiming is the better of the two approaches and has fewer faults. Thus, we will be interested in repeaters that are self-timed and employ forward retiming.

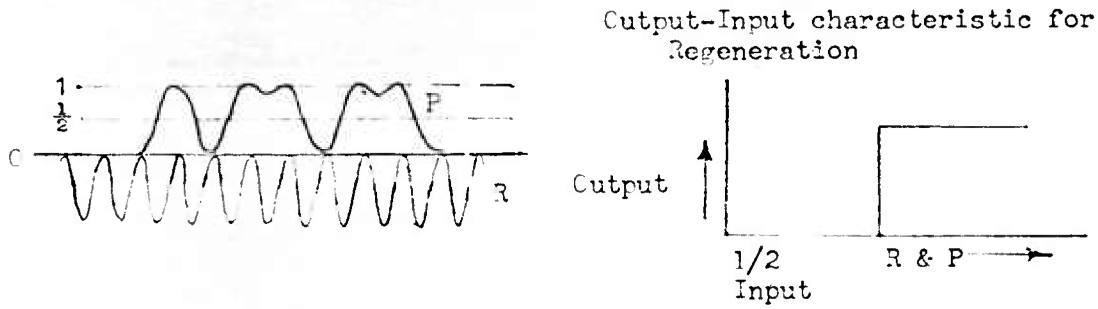
(4) Once the timing signal is obtained, it is used in detection and to control pulse location when regenerating. Three methods of employing the timing are listed below:*

(a) Partial retiming

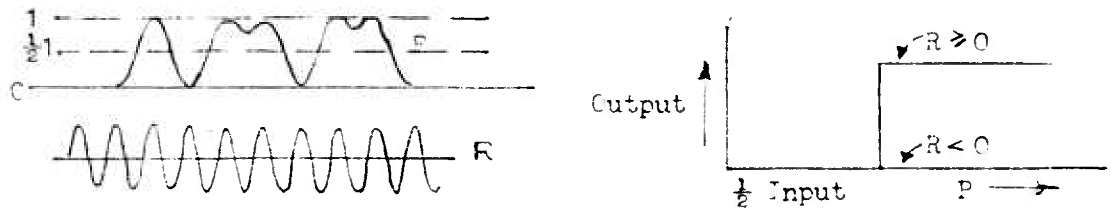
(b) Permissive retiming

(c) Complete retiming.

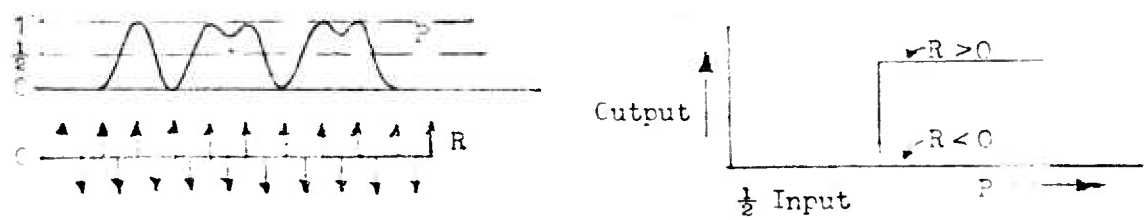
Each of these is shown in Figure II.



(a) Partial Retiming



(b) Permissive Timing



(c) Complete Retiming

FIGURE II. Partial, Permissive and Complete Retiming

*Page 105, ref 3.

(5) Partial retiming adds the timing signal directly to the amplitude pulse train. The resulting sum is threshold detected for "1's" or "0's." Permissive retiming uses the timing signal in an "and" gate action to sample the amplitude pulse train. When the timing signal exceeds a certain threshold, the pulses out of the preamplifier are examined to see if they are "1's" or "0's." Finally, complete retiming employs only the zero crossings of the timing signal. At a positive going cross-over, a narrow sampling pulse is generated. If the amplitude of the pulse train being detected exceeds the threshold at that instant, a "one" is generated and if it does not exceed the threshold, a "zero" is generated. The particular level that is generated continues until the negative going cross-over of the timing wave turns it off.

(6) Each of these retiming approaches is discussed in detail in Reference No. 3. Complete retiming, although generally the most difficult to implement, seems to offer the best tolerance to noise and other system imperfections. We will now describe a few of the practical problems associated with retiming.

3. Practical Problems Associated with Retiming

a. Timing Recovery

(1) Up to this point, we have indicated different approaches that can be used for timing extraction (forward, backward, etc.). However, we have not mentioned how the timing signal is physically removed from the digital train. If the pulse train contains a periodic component at the clock frequency, a narrow band filter is all that is required to extract timing. If this is not the case, other processes, such as rectification, may be required before filtration.

(2) Generally there are three different types of narrow band-pass filters employed.

(a) L - C filter

(b) Crystal filter

(c) Phase-lock oscillator

(3) The use of all three types in digital repeaters has received extensive treatment.[1,4] The problem of which one to use in a given application becomes a matter of weighing cost and complexity against circuit stability and reliability. Narrow band L - C filters have a greater variation in center frequency with temperature than crystal filters or phase-lock oscillators and a practical limit to their upper value of Q.

(4) Crystal filters have a practical higher upper value of Q, but are still temperature sensitive. Phase-lock oscillators, if properly designed, are very stable and approach infinite Q. While the phase-lock oscillator seems to possess good characteristics, it is the most complex of the three, while the crystal filter would be physically the largest.

Regardless of which filter is chosen, each has certain inherent defects that contribute to timing error. The next section of this report deals with a variety of sources of timing error.

b. Sources of Timing Error

There are many sources of timing error which contribute to timing jitter. This error can cause misplaced sampling in detection and a variation in the leading edges of the regenerated pulses. Timing jitter increases the chance of falsely detecting "1's" and "0's" and thus increases the probability of error in transmission. A few of the contributors to timing error will now be discussed in general terms.

(1) Intersymbol Interference

Intersymbol interference arises when transmitting pulses over a band limited channel. This occurs when the higher frequency components of a signal are removed causin a pulse to spread into adjacent intervals. This interference may be increased if the transmission medium is not linear phase, thus delaying some frequency components more or less than others. Intersymbol interference, which amounts to the overlap of adjacent pulses, can be a very great source of error. The amount of overlap varies and depends upon the received pulse patterns. It not only affects the extracted timing, but influences threshold detection as well. In order to reduce its effects, we attempt to equalize the transmission medium's phase so it becomes as linear as possible. Also, the effect on the extracted timing signal can be reduced if high Q filters are employed to average the variations with pulse patterns.

(2) Noise and Crosstalk

(a) Another common source of timing error is produced when a transmitted pulse is subjected to outside interference. The most common types are Gaussian noise, shot noise, cable cross-talk and other forms of cable pickup (i.e., radio-frequency interference).

(b) All of these tend to affect threshold detection; however, of the above four, cable cross-talk is the greatest source of timing error. Cable cross-talk is generated by induction from an adjacent cable carrying pulses at the same bit rate as the transmitted signal. When extracting timing with a high Q filter, shot and Gaussian noise are not a serious problem due to their broad frequency spectrums. However, cross-talk has its energy concentrated in the band pass of the filter and will appear in the extracted timing waveform. This is also true of any cable pickup with a frequency at the transmission rate. The only method of reducing this source of interference is removal of the source or some sort of shielded balanced cable.

(3) Detuning

External interference and cable characteristics are not the only sources of timing error. Proper filtration and recovery of the timing signal require tuning the band-pass filter to the transmitted periodic component.

If this is not the case, two types of error or jitter can result. The first, known as steady-state jitter, results in the extracted component being phase shifted a fixed amount relative to the original data. The second, called dynamic phase jitter, results in an ever-changing phase shift that depends on the received pulse pattern. The cause of detuning can be from either of two sources. Either the crystal producing the clock frequency has drifted from its initial adjustment, or the center frequency of the band-pass filter has changed due to aging or temperature variation. In both cases, the results can be very harmful if not controlled. Steady-state jitter becomes appreciable in a high Q filter, since any variation from the tuned frequency of the filter will cause a large shift in the filtered signal. The effect of this phase shift, within a given repeater, is improper location of sampling for detection. Its effects, however, are not accumulative from one repeater to the next. Dynamic phase jitter offers a more severe problem because its effects are accumulative over a repeatered system. It arises in a PCM type system where pulses are present only part of the time. A filter will have a steady-state response which is a function of the driving source, as well as a transient response which depends only on the circuit configuration and initial conditions. In a high Q filter, used to extract timing from a pulse train that does not always have pulses present, we require the natural frequency of the filter (transient response) to be the same as the frequency of the driving signal (steady-state response.) If this is not the case, the filter will extract one frequency when pulses are present, and its frequency will change when the source is temporarily withdrawn. In some PCM systems it is possible to go many intervals without any pulses present and we need to rely on the natural ringing of the filter for timing. Differences between natural and source frequencies are reflected in variation in phase that depend on how frequently transmitted pulses arrive. This is dynamic phase jitter. Its accumulative effect, which causes variations in the leading edges of the regenerated pulses, can cause a system to fail after a number of repeatered links. We may overcome these difficulties by removing detuning effects if possible, or by choosing transmission modulation approaches which are not susceptible to this form of error.

(4) Nonlinear Effects

The most important sources of error have been discussed, but there are some others that are worth mentioning. One of these is nonlinear effects. In extracting timing, it may be necessary to carry out a nonlinear operation such as rectification of the received digital pulse train. Practical rectification may not be ideal, thus introducing additional error. There are also other effects such as pulsing a filter with finite width pulses that may add error but generally cannot be avoided.

(5) Conclusion

The primary sources of timing error are crosstalk, detuning, and intersymbol interference. Timing error can cause improper detections and since this error can be cumulative, it limits the practical length of the PCM type system. The method of reducing timing error involves the use of specialized circuitry and improved transmission modulation approaches.

4. Modulation Approaches

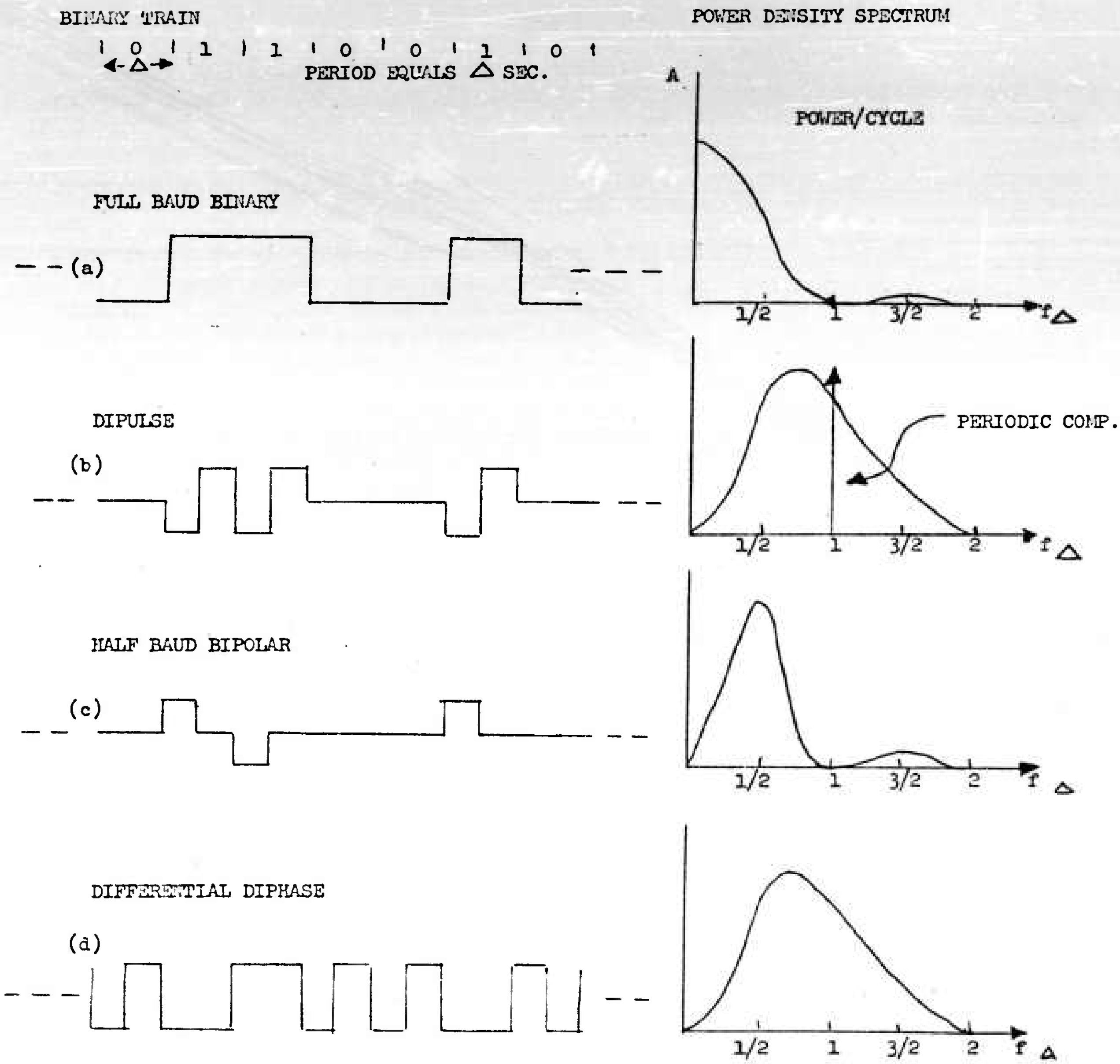
There are a number of ways of modulating PCM type signal for transmission over a cable. We will discuss a few of these briefly. Of particular interest are full baud binary, dipulse, bipolar and differential diphase.

a. Full Baud Binary

Full baud binary involves sending an up level when a one is present and a down level when a zero is present. Referring to its power density spectrum (FIG. IIIa), it extends down to dc and does not possess a periodic component. One of the serious disadvantages associated with full baud binary is the fact that it has a lot of power at the low frequencies. In military PCM systems, a voice frequency order wire is required and is transmitted along with the PCM traffic. For this reason, it is wise to choose a modulation approach which does not have energy down near dc. Also, the use of transformers will introduce high pass filter effects which cause signal droop and severely affect detection of full baud binary.

b. Dipulse

Dipulse amounts to sending a full cycle of a square wave for a one and nothing for a zero (FIG. IIIb.) Its power density spectrum contains a periodic component at the Pulse Repetition Rate (PRR). In addition, its power spectrum is zero at dc. This is important, and both factors make dipulse a useful modulation technique. The periodic component can be extracted using an ordinary filter within a regenerative repeater. However, it is theoretically possible for long strings of zeros to occur, during which period nothing is sent. This introduces timing error as discussed earlier. Also, the power spectrum is twice the width of full baud binary and bipolar. The wider bandwidth increases signal loss in transmitting it over a low pass network.



SIGNAL WAVE FORMS AND POWER DENSITY SPECTRUMS

FIGURE III

c. Bipolar

Figure IIIc demonstrates $\frac{1}{2}$ baud bipolar presently being used by Bell Telephone Laboratories.[3,4] Its power spectrum does not have much low frequency power or a periodic component. Alternate ones are sent with opposite polarity pulses and nothing is sent for a zero. In order to extract a periodic component, rectification of the received signal is required before filtration. Bipolar is useful since its energy is concentrated at lower frequencies than dipulse and thus suffers less energy loss over cable. In addition, in the Bell Labs application, crosstalk is a prime problem due to multi-pair cable in close proximity. Bipolar has a lower crosstalk figure than dipulse when used with the balanced cable Bell Labs employ. The possibility of sending long strings of zeros still exists, but is alleviated by proper coding. The primary influence in using bipolar is the reduction of crosstalk, but dynamic phase jitter still exists since periods of no pulses can still occur.

d. Differential Diphase

Differential diphase is shown in Fig IIId. When a zero is present, no phase reversal occurs; while if a one is present, a phase reversal occurs. This amounts to sending a full cycle square wave for zero and a half cycle for a one. Its power spectrum* is the same as dipulse, but does not have a periodic component. The various advantages of differential diphase are not apparent from the power spectrum alone. Generally, a differentially coherent detection system is required. However, a new method of detection, to be discussed later, offers many advantages over other past diphase detection procedures. The remainder of this report is devoted to a detailed discussion of a differential diphase system.

5. Conclusion

This concludes the background material for this report. We have outlined the problems associated with digital transmission over a cable. A few common modulation techniques were briefly discussed. We will now demonstrate how differential diphase can be used in repeatered cable transmission system.

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*Calculated in Appendix I

DIFFERENTIAL DIPHASE REPEATERED SYSTEMS

1. Diphase Transmission and Detection

a. Differential diphase is a differentially coherent signaling technique and its detection should depend on changes in phase. O. Ringelhaan,[5] of Communications Security Division, USASRDL, developed a method of eliminating direct use of phase information in detection, and thus reduced the procedure to ordinary threshold detection.

b. Fig. IV describes the basic steps in detection and regeneration. Line (b) represents the original diphase before transmission over one mile of CX-4245 cable. (PRR is 2.304 megabits/sec.) If the received signal is only phase equalized at 1.152 and 2.304 megacycles, the signal shown in line c results. The high frequency portion or zeros are more severely attenuated than the lower frequency ones since no amplitude equalization is employed. If the phase equalized signal is rectified, the resulting wave form, line d, will have a periodic component. The 2nd harmonic of the larger lobes and the 1st harmonic of the smaller lobes contain a 4.6-Mc signal. When the periodic component is removed from the rectified signal, the resulting pulses resemble the triangular wave forms shown on line "f." Ordinary threshold detection can then be employed to identify ones and zeros. Using the timing signal of line "e" and the detected signal of line "f," we can retime and regenerate the differential diphase wave form. One method of accomplishing regeneration is by using squared up versions of lines "e" and "f" in an "or" gate action and then sending the result into a complimenting flip-flop. With this de. ription of diphase reconstruction, we can begin our analysis of diphase signaling.

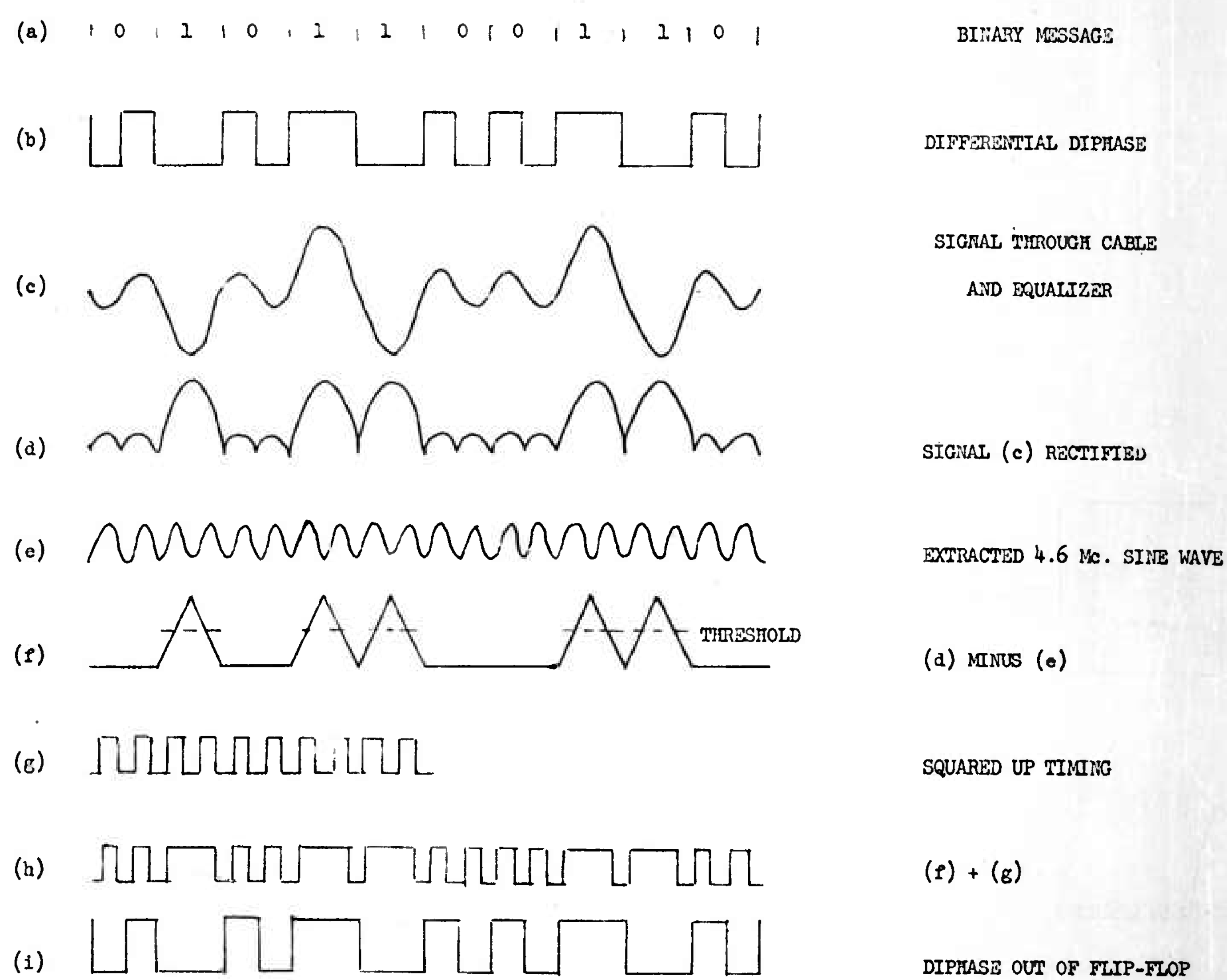
2. Analysis of Differential Diphase Signaling

In investigating differential diphase, we find a number of topics to be considered. Among those that will be discussed are retiming, detectability, and regeneration.

a. Retiming in a Diphase Repeater

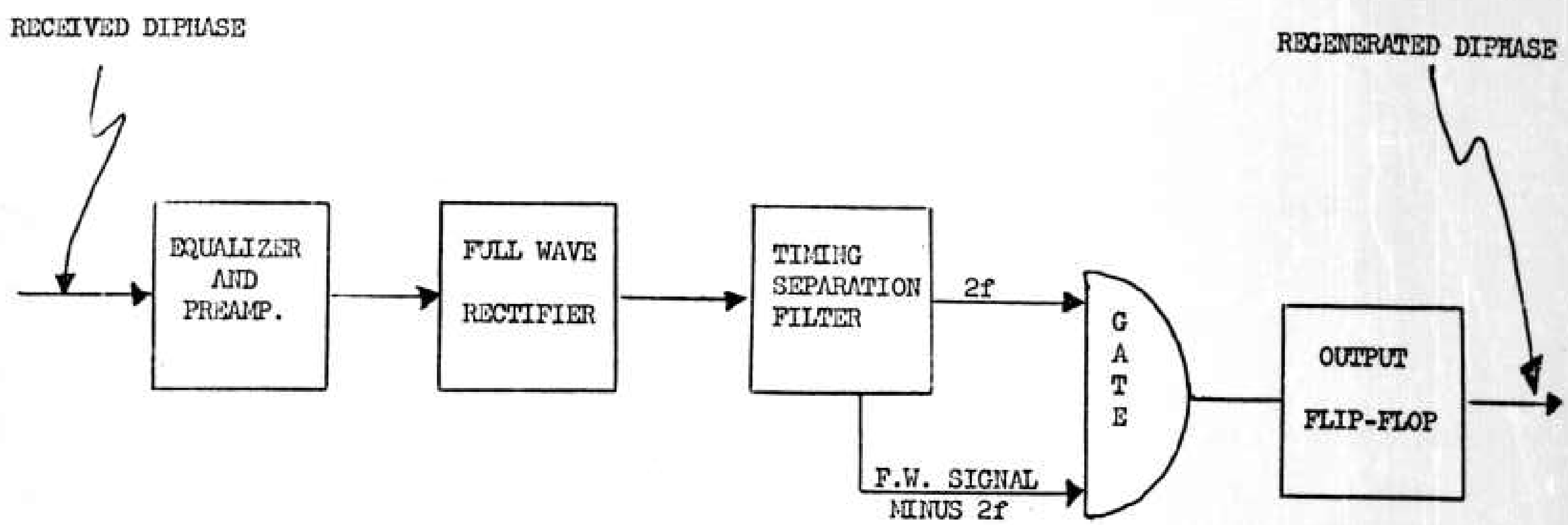
(1) A block diagram of the initially proposed diphase repeater is shown in Fig. V. It essentially carries out the operations described in Fig. IV.

(2) If we were to examine a typical differential diphase signal after it passed over a mile of CX-4245 cable, it would be very difficult to distinguish 1's from 0's because of a loss of zero crossing information. In addition, a number of differently phased pulses would result instead of just a symbol for a one and another for a zero. This effect arises primarily because the cable is not an ideal transmission line at the frequencies of concern and thus the phase characteristic is not linear. If the phase characteristic can be made reasonably linear, the cross-over points correspond to those of the original signal. The former square waves then resemble sine waves at either 1.15 megacycles or 2.3 megacycles. A simple phase equalizer which is employed is a loosely coupled L-C filter tuned to 2.3 megacycles. The corrective effects of such a network are sufficient for this application.



REPEATER MODE OF OPERATION

FIGURE IV



BLOCK DIAGRAM OF PART OF DIPHASE REPEATER

FIGURE V

After rectification and filtration we obtain a 4.6-Mc signal as described earlier. Since both the ones and zeros contribute to this timing signal, a problem peculiar to this system results.

(3) Ideally speaking, we would like the 4.6-Mc component contained in both the rectified ones and zeros to be equal in both phase and amplitude. If this condition exists, a continuous timing signal is present and a rather simple filter is required for its extraction. To achieve this situation, two requirements must be met:

(a) The amplitude ratio of ones to zeros should be 5:1 since the 2nd harmonic of the ones is 1/5 the magnitude of the 1st harmonic of the zeros.

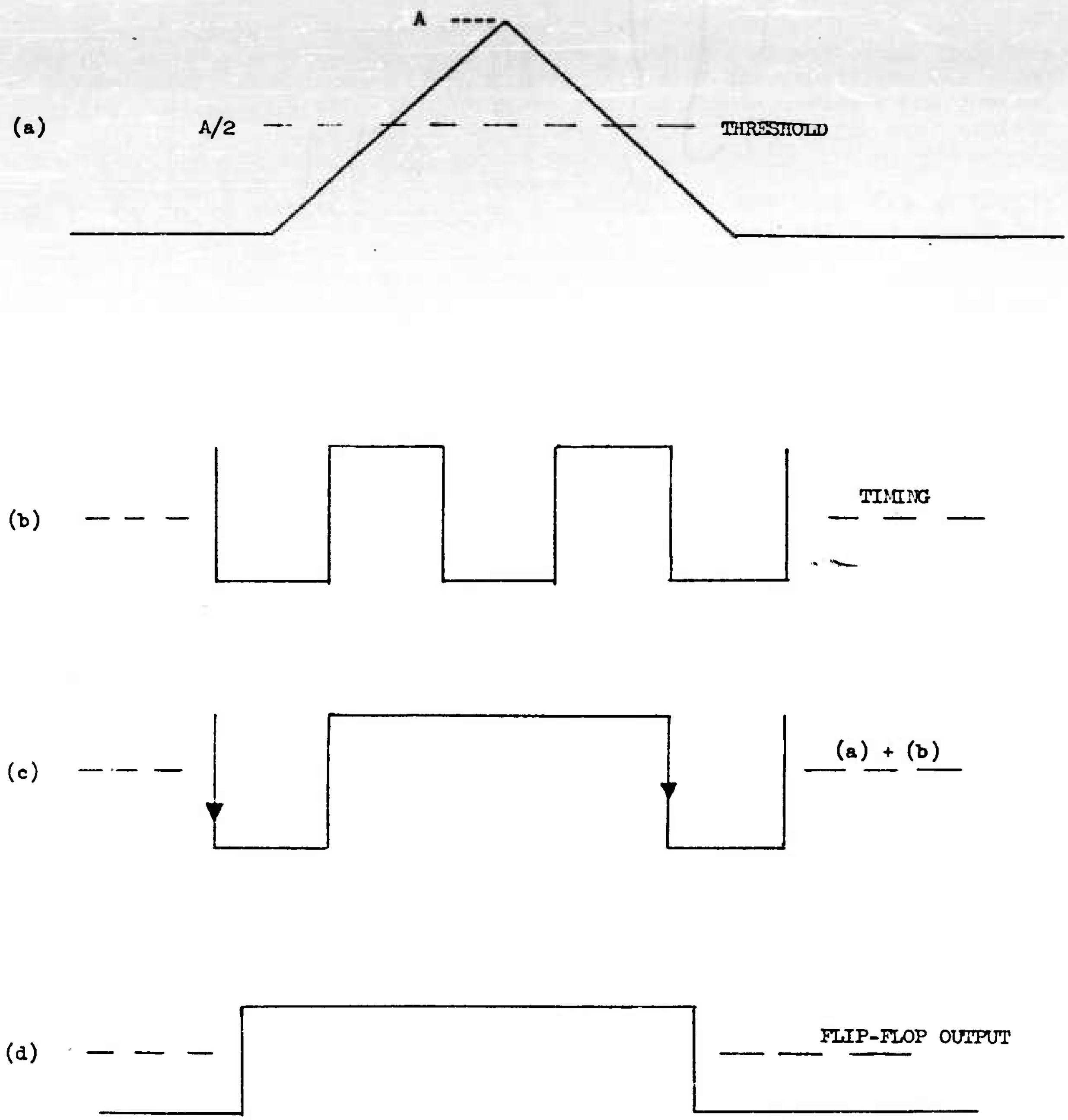
(b) The phase delay of the zeros (2.3-Mc signal) and ones (1.15-Mc signal) should be linear so that after rectification, their 4.6-Mc components are in phase.

(4) If the 2nd of the two conditions is not met, dynamic variations in the phase of the timing signal can result. Such a condition might occur if after the equalizer is designed, the cable or various component characteristics vary. Provided the Q of the timing filter is reasonably high, minor variation will not be bothersome. However, the same cannot be said if variations become too great. A laboratory measurement was made with Q's of roughly 100 and variation in cable length of up to $\pm 10\%$. Very little dynamic phase variation was observed. The avenue to more conclusive results in this area, we feel, is through measurement rather than computation since the problem depends upon a variety of factors.

(5) The timing error which results when the received amplitude ratios of 1's and 0's are not exactly 5:1 is minor. The 5:1 ratio in amplitude can be maintained roughly by proper adjustment of the input impedance of the equalizer.

(6) After the timing signal is obtained, it is utilized in the detection procedure. A detailed description of the operations is shown in lines "f" through "i." An enlarged version of lines "f" through "i" is shown in Fig. VI. By a careful examination of Fig. VI, the tolerable phase variations of the timing signal can be determined. Initially, a noise-free signal will be assumed.

(7) If the timing train is in its correct position, it accomplishes regeneration as shown in Fig. VI. When the timing is up, the ORed train must be up regardless of the amplitude of VI(a). The only time the amplitude level of VI(a) determines the ORed signal level is when the timing signal is down. Therefore, we can think of the down timing level in the middle of the triangular pulse, as a sampling window. The detected signal will be above its threshold if a one is present and below its threshold if a zero is present. This determination is passed on to the combined train VI(c). Train (c) is sent to a flip-flop and this generates the actual diphase. Therefore, correct regeneration of the differential diphase signal depends upon obtaining waveform VI(c) correctly. Since the negative shifts of line (c) are used to trigger the output flip-flop, errors can result only when false negative shifts are present or true negative shifts fail to occur. Our analysis then will consist of determining how much phase variation in timing is



CORRECT DIPHASE REGENERATION

FIGURE VI

possible before the above improper shifts can occur.

(8) We will consider the two extremes of phase shift in timing, first to the left and then to the right. Figure VII shows the timing shifted $\pm 90^{\circ}$ from its original location.

(9) If the timing is shifted more than 90° to the left, errors can result. A negative shift will occur in the ORed train just before point E that would not have occurred had the timing location been correct. This false negative shift will improperly toggle the flip-flop causing an error. At point F, if the timing wave is more than 90° left of its correct position, a negative shift whose location depends on the amplitude level of train (a) rather than the timing signal will occur. It is harmful to have VI(a) determine the location of negative shifts since it contains more noise than the filtered timing. Therefore, the timing should not exceed 90° phase shift to the left.

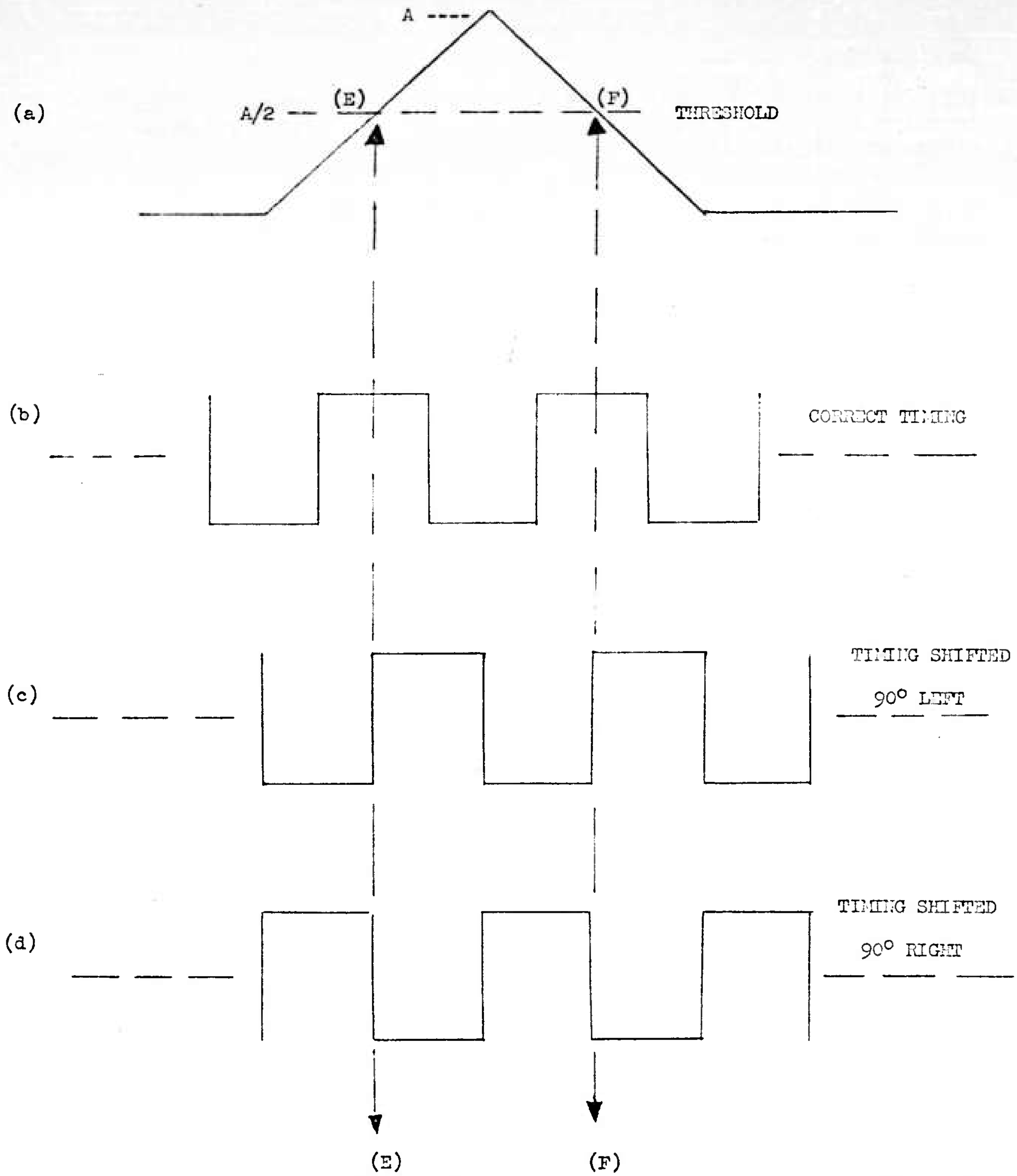
(10) By a similar analysis we can show that the timing should not exceed 90° phase shift to the right. Our total tolerable variation in timing phase is $\pm 90^{\circ}$. Practically speaking, this figure is closer to $\pm 80^{\circ}$ due to circuit inaccuracies. This phase figure is considered good in comparison with other modulation approaches such as dipulse. This improvement arises because the timing signal, used to sample VI(a), is four times the pulse repetition rate of VI(a). Therefore, large phase deviations in timing amount to only small phase changes relative to the detected amplitude waveform. Here, 4° phase variation relative to the timing signal will amount to only 1° variation relative to the sampled wave.

(11) The effects of phase variations within the tolerable $\pm 80^{\circ}$ cause no difficulties in detection unless the detected amplitude signal contains noise. If the amplitude signal contains noise, then as we move the sampling window from its correct location, the probability of false detection increases.

(12) This leads us to the problem of a diphase signal that is contaminated with noise. The effects of noise on the diphase signal are felt in the extracted timing and sampled amplitude waveform. Noise in the timing signal will produce phase variations, the effects of which have just been discussed. The effect of noise in the amplitude waveform is discussed below.

b. Detection in a Diphase Repeater

(1) In order to reduce the probability of making an error when using detection, there are certain obvious rules we should follow. Some of these include (a) Sampling at a point where the signal mean separation between a one and a zero is greatest; (b) Using as narrow a sampling window as possible; and (c) Using complete retiming in generating the sampling window.



SHIFTED TIMING

FIGURE VII

(2) We can now make some objective comments about the methods O. Ringlehaan employed in his repeater. The width of the sampling window is fairly narrow but may be made narrower and less susceptible to jitter by employing the derivative of the timing wave. The "OR" gate action used in sampling the amplitude signal is a poor feature. If at any instant when the amplitude signal is being sampled, a supposed one falls below the threshold or a supposed zero exceeds the threshold, an error in detection will result. This causes the system to be very susceptible to errors produced by noise. This problem can be partially solved by modifying gating action somewhat. One improvement consists of detecting a one if the amplitude exceeds its threshold during any instant within the sampling interval. This will improve our chances of detecting ones correctly without affecting the detection of zeros. There are other possibilities for improvement of the sampling process that future investigations should unfold.

3. Comparison between Differential Diphase and Dipulse Modulation

Diphase and dipulse signaling will be compared with reference to the various sources of error mentioned in the background section. These include noise, crosstalk, and detuning.

a. Effects of Error Source on Timing in Diphase and Dipulse

(1) Noise, crosstalk and detuning all contribute to phase error or phase jitter. For the moment we will neglect crosstalk and consider white noise and detuning.

(2) Extensive work has been done in analyzing the effect of noise and detuning on timing jitter in digital repeaters.[6,7] O. E. DeLange points out that one source of cumulative jitter is caused by detuning.

This jitter is characterized by the following expression:

$$\phi\ ms = \frac{K_1\ N^{\frac{1}{2}}}{Q} \cdot \quad W_o + N^2 K_2\ Q \quad (1)$$

ϕ = Phase jitter, K_1 = const depending on system, N = number of repeaters, Q = filter Q, W_o = random noise of the signal and additional white noise, K_2 = const depending on detuning. The first term is jitter contributed by noise while the second is phase jitter due to detuning. It is possible to find an optimum Q by differentiating expression (1) with respect to Q, and setting the result equal to zero. Equation (1) was derived for a signal with an average 50% duty cycle. The second term occurs because the timing signal is derived from pulses that are present only part of the time. This situation occurs when dipulse signaling is employed. However, when differential diphase modulation is used, the timing signal is present all the time and the second term is equal to zero. This is an extremely useful characteristic in diphase. The resulting dynamic phase jitter depends only on noise and not on detuning. The larger the Q, the less the phase jitter. One of the primary drawbacks to dipulse was the accumulation of phase jitter due to the detuning in a multi-repeatered system. Here, this problem is not as serious.

(3) Thus, detuning does not affect dynamic phase jitter in a diphase repeater. The only limiting factor in determining a value for tolerable detuning is steady-state phase jitter. In a diphase repeater this should not exceed $\pm 80^{\circ}$. An ordinary L-C filter with a Q of 100 may offer adequate stability and noise filtration for our application. The design aspects of the filter are not considered here.

(4) There are also other advantages in employing diphase signaling. In a dipulse repeater, it is possible for several zeros to be transmitted in succession causing a long period of no signal. The Q of filter must be high so that it rings during the entire period that zeros are sent. In a diphase repeater this situation will not occur and a lower Q filter is sufficient.

(5) Crosstalk from an adjacent cable can also contribute to phase error. When dipulse is used, the effects of crosstalk on timing are most pronounced when zeros are transmitted on the line. This is due to the fact that crosstalk is the sole influencing effect on the timing filter when zeros are received. When diphase is employed, this situation will not occur since the signal will always be present.

(6) From this brief commentary it is clear that differential diphase offers a great improvement over dipulse in retiming. By removing most of the cumulative phase error, the number of unattended repeaters in a series may be increased without a marked increase in the error probability (P_e).

b. Detectability of Diphase and Dipulse

Aside from the improvement in retiming, diphase offers an advantage in detectability. In order to demonstrate this effect, the curves of the probability of error vs peak signal to rms noise are plotted for both diphase and dipulse in Fig. VIII. Gaussian noise is assumed. The details of the computation are shown in Appendix II. At a probability of error equal to 10^{-6} a saving of 10 db in S/N is obtained. We must keep in mind that a number of idealizations were made in obtaining these figures but they do indicate that the diphase offers a definite improvement in peak signal to rms noise threshold.

c. Practical Considerations

The initial diphase repeater model proposed by O. Ringelhaan appears to offer some practical advantages over dipulse. L-C rather than crystal filter is employed and thus the expected cost is less. More detailed treatment of this subject will be contained in a future report by M. Hooten.

CONCLUSIONS

1. The differential diphase modulation approach offers a number of theoretical and practical advantages over dipulse modulation. These include improved retiming, better detectability in noise, resistance to crosstalk,

and less expensive repeaters. The results presented in this report, however, are far from conclusive.

2. We have considered only a single repeater and as yet have not tested a multi-repeatered system. The cumulative effects of improper equalization have not been thoroughly determined and also final design of the repeater is not fully completed. From this preliminary investigation, it appears that diphase modulation offers a sound starting point for a digital transmission system. Final results will only be known after a multi-repeatered system has been built and tested.

RECOMMENDATIONS

1. Recommendations for improving the diphase repeater:

   a. Modifying the gate action in regeneration.

   b. Narrowing the sample window.

   c. Incorporation of AGC, if necessary.

   d. Employing more sophisticated regeneration in the attended repeaters.

2. Areas of future investigation should include:

   a. More extensive testing concerning phase equalization.

   b. Determination of the necessary filter Q in the repeater.

   c. Total system tests involving many repeaters.

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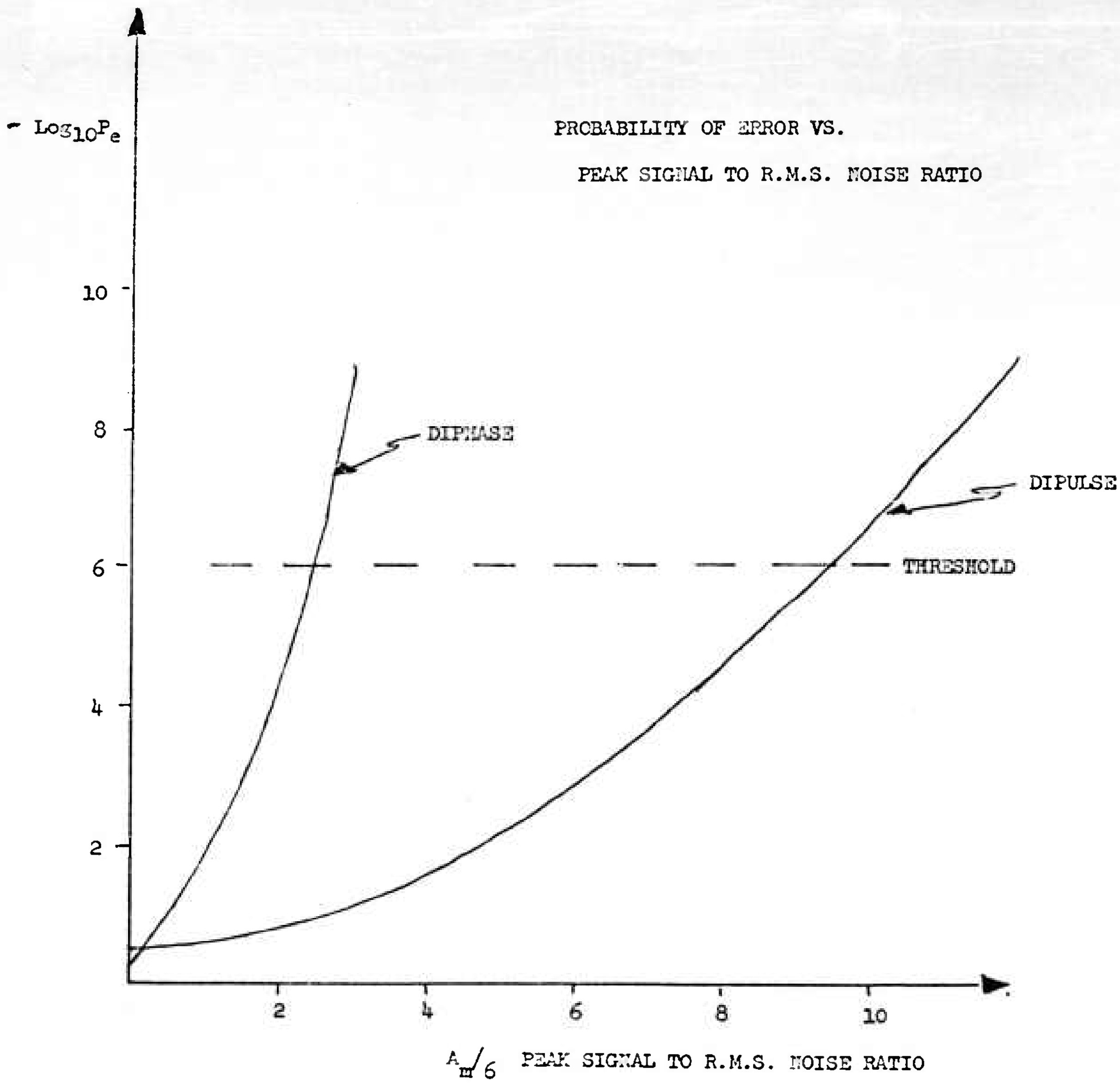


- Log10Pe
PROBABILITY OF ERROR VS.
PEAK SIGNAL TO R.M.S. NOISE RATIO
10
8
6
4
2
DIPHASE
DIPULSE
THRESHOLD
2
4
6
8
10
Am/6 PEAK SIGNAL TO R.M.S. NOISE RATIO

FIGURE VIII

APPENDIX I
CALCULATION OF POWER DENSITY SPECTRUM

The calculation of the power density spectrum for differential diphase will be carried out. A number of different methods can be employed, however, when a rectangular PCM train is considered, simplifications in the computation are possible.

Initially, we will find the autocorrelation and then take its fourier transform to get the desired power density spectrum. The binary train will be assumed to consist of ones and zeros occurring in a random manner each independent of the other with equal probability equal to 1/2. A full-cycle square wave is sent for a zero and a 1/2 cycle for a one. A typical signal is shown in Fig. III(d).

When dealing with PCM pulse trains that are composed of rectangular shaped pulses and of known duration, one can show the resulting autocorrelations are composed of a series of continuous straight line segments. Thus, one needs only to determine the value of the autocorrelation at the points where the line segments come together and the total correlation is determined. The straight line segments have changes in slope at intervals related to the duration of the pulses composing the PCM train. With this brief commentary, we will commence with finding the autocorrelation of diphase.

The expression for finding the autocorrelation of a stationary discrete process is obtained by averaging across the ensemble at two discrete points separated by a distance (T) and is given below.[8]

$$\phi_{11}(T) = \sum_{j=-\infty}^{\infty} \sum_{i=-\infty}^{\infty} x_{1i} \; x_{2j} \quad P(x_{1i}, x_{2j}; T)$$

x_{1i} is the value of the first observation and x_{2j} is the value of the second, while $P(x_{1i}, x_{2j}; T)$ is the joint probability density of the occurrence of x_{1i} and x_{2j} separated by T.

All we need find is the $\phi_{11}(T)$ for T in multiples of $n\Delta$*/2 since these points determine where the straight line segments of $\phi_{11}(T)$ come together. Initially, we will consider $\phi_{11}(0)$.

$$\phi_{11}(0) = \sum_{j=-\infty}^{\infty} \sum_{i=-\infty}^{\infty} x_{1i}, x_{2j} \quad P(x_{1i}, x_{2j}; 0).$$

The only possible values for the x's are 0 and E. Products involving x_1 or x=0 will not contribute to the correlation. Thus, only terms where $x_1=x_2=E$ will contribute. All we need find is:

$$\phi_{11}(0) = E^2 \sum_{j=-\infty}^{\infty} \sum_{i=-\infty}^{\infty} P(E_{1i}, E_{2j}; 0)$$

E is value of signal when an up occurs.

$\sum_{j=-\infty}^{\infty} \sum_{i=-\infty}^{\infty} P(E_{1i}, E_{2j}; 0)$ = Probability of finding the value E at time 1.

$$\phi_{11}(0) = \frac{E^2}{2}.$$

* Δ is baud width

Next we will find $\phi_{11}\left(\frac{\Delta}{2}\right)$

(1)
$$\phi_{11}\left(\frac{\Delta}{2}\right) = E^2 \sum_j \sum_i P\left(E_{1i}, E_{2j}; \frac{\Delta}{2}\right); \quad \sum_j \sum_i P\left(E_{1i}, E_{2j}; \frac{\Delta}{2}\right) =$$

(Probability of finding an E at time t_1) (Probability of finding an E at t_2/given an E at t_1) where t_1 and t_2 are separated by $T = \frac{\Delta}{2}$, =

$\frac{1}{2} \times \left(\frac{1}{4}\right) = 1/8$

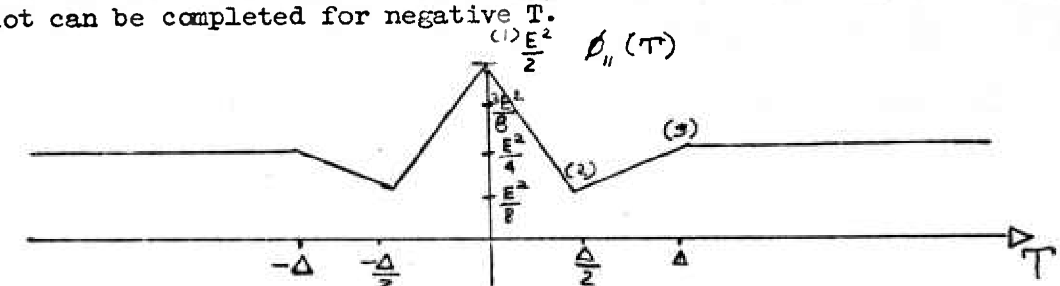
(2) $\therefore \phi_{11}\left(\frac{\Delta}{2}\right) = \frac{E^2}{8}$.

In a like manner we can show

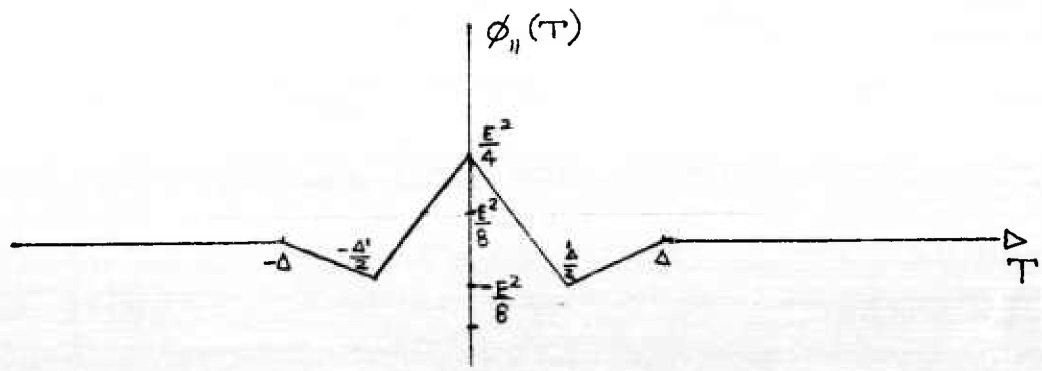
(3) $\phi_{11}(\Delta) = \frac{E^2}{4}$.

(4) Also $\phi_{11}\left(n\frac{\Delta}{2}\right) = \frac{E^2}{4} \quad n \geq 2$.

These points can be plotted and the straight line segments filled in. If we use the fact that autocorrelations are symmetrical about their origin, the plot can be completed for negative T.



If we assume the signal has no dc component by allowing it to swing from $\frac{+E}{2}$ to $\frac{-E}{2}$ the autocorrelation plot loses its dc term as shown below.

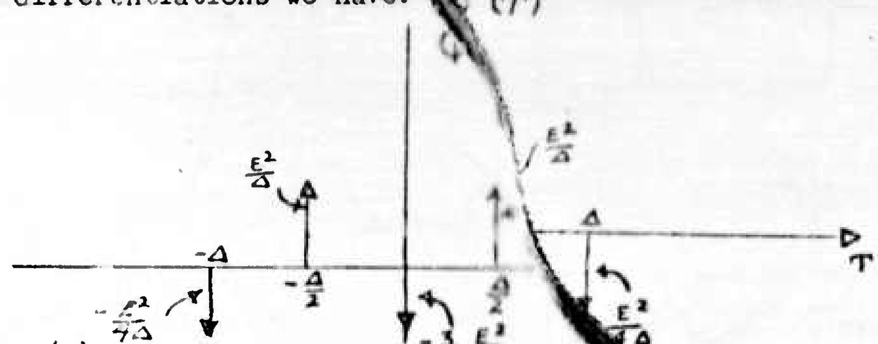


If we take the fourier transform of ϕ_{11} (T) we have the desired plot in Fig. IIId.

$$\phi_{11}(w) = \frac{1}{2\pi} \times \int_{-\infty}^{\infty} \phi_{11}(T)\, e^{-jwT}\, dT.$$

By using short cuts suggested by Mason and Zimmerman, this process is made simple by successive differentiation of $\phi_{11}(T)$ until the plot reduces to a series of impulses. The fourier transform of an impulse is merely a constant with an $e^{\pm jwa}$ term attached if it is displaced "a" units from the origin. Also, using the fact that $f'(t) = jwf(w)$, we can find the desired transform of $\phi_{11}(T)$.

After two differentiations we have: $\phi''_{11}(T)$



$$\phi''_{11}(T) = -w^2\, \phi_{11}(w)$$

$$-w^2\, \phi_{11}(w) = \frac{-1}{2\pi}\,\frac{E^2}{4\Delta}\left(e^{+jw\Delta} + e^{-jw\Delta}\right) + \frac{1}{2\pi}\,\frac{E^2}{\Delta}\left(e^{+jw\frac{\Delta}{2}} + e^{-jw\frac{\Delta}{2}}\right) - \frac{1}{2\pi}\,\frac{3}{2}\,\frac{E^2}{\Delta}$$

$$\phi_{11}(w) = \frac{1}{2\pi}\,\frac{E^2}{\Delta w^2}\left(-2\cos w\frac{\Delta}{2} + \frac{1}{2}\cos w\Delta + \frac{3}{2}\right)$$

$$\phi_{11}(w) = \frac{E^2}{4\pi\Delta w^2}\left(3 - 4\cos w\frac{\Delta}{2} + \cos w\Delta\right)$$

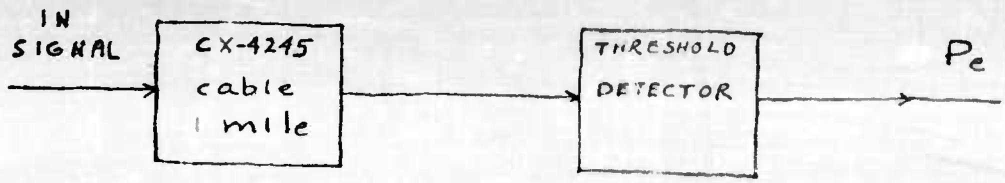
$$\phi_{11}(w) = \frac{E^2}{2\pi\,\Delta\, w^2}\left(1 - \cos w\frac{\Delta}{2}\right)^2.$$

This is plotted in Fig. IIId as a function of $f\Delta$. All the other plots in Fig. III can be found in a similar fashion.

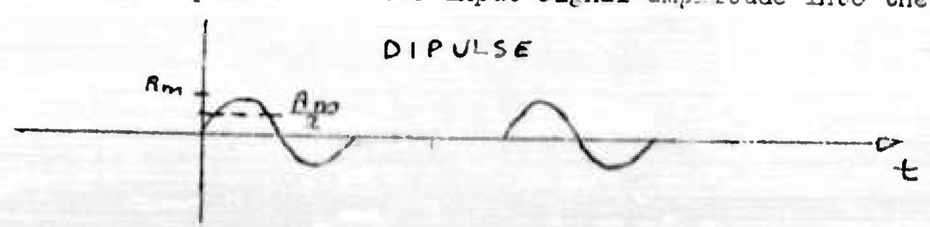
APPENDIX II
CALCULATION OF PROBABILITY OF ERROR VS PEAK SIGNAL TO RMS NOISE RATIO

This section is concerned with calculating the probability of error vs. peak signal to rms noise ratio for diphase and dipulse signaling. Additive Gaussian noise is assumed along with ideal threshold detection. This calculation should give a general idea as to which approach offers a greater resistance to noise. However, since we have not considered the actual detection systems involved, the results are not complete. A truer comparison of the two systems will be found from experimental measurements.

The following system may be assumed for our computation.



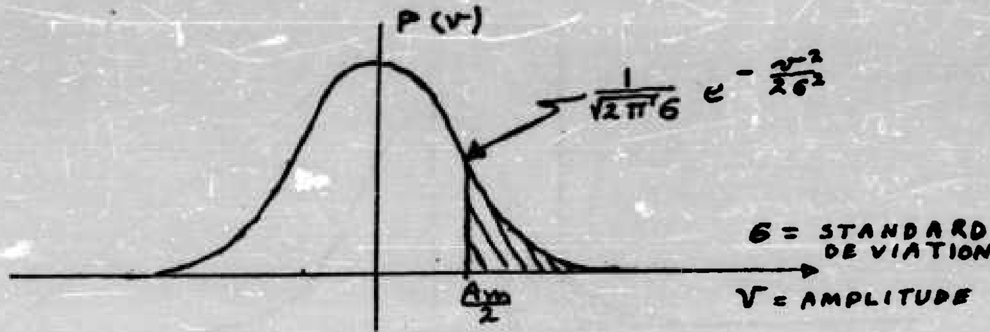
In calculating the peak signal to rms noise we will fix the amplitude of the "in signal" and compare the systems on this basis. Both the cable characteristics and detector play a part in determining the total desired plot. In passing over the cable, the dipulse ones are attenuated 12 db more than the diphase ones and the same as diphase zeros. This is due to the fact that the fundamental frequency of the diphase ones is half the fundamental frequency of dipulse ones and diphase zeros. At the far end, diphase is rectified before threshold detection occurs. It will be shown that rectification increases the probability of error by 3/2 assuming additive Gaussian noise. Both diphase and dipulse are then subjected to threshold detection with the threshold located halfway between zero and the signal peak voltage. First, we will calculate the Pe vs. peak signal to rms noise ratio for dipulse. Let the input signal amplitude into the detector = A_m.



In order to find the probability of making an error we have:
P (error) = P (detecting a one) P (a zero was present/given we detected a one) + P (detecting a zero) P (a one is present/given we detected a zero.)

We can determine the conditional probabilities by examining the amplitude distribution of the Gaussian noise.

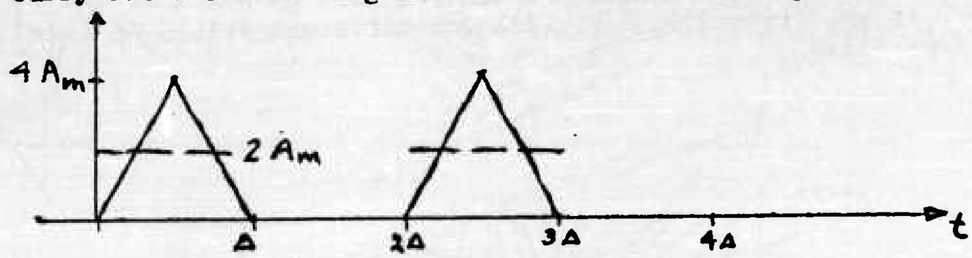
P (a zero is present/given we detected a one) = P (a one is present/given we detected a zero) = shaded area.



$$\therefore P_e = \frac{1}{2}\int_{\frac{A_m}{2}}^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma}\, e^{-\frac{v^2}{2\sigma^2}}\, dv + \frac{1}{2}\int_{\frac{A_m}{2}}^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma}\, e^{-\frac{v^2}{2\sigma^2}}\, dv$$

$$P_e = \int_{\frac{A_m}{2}}^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma}\, e^{-\frac{v^2}{2\sigma^2}}\, dv$$

We can carry out a similar argument for differential diphase.

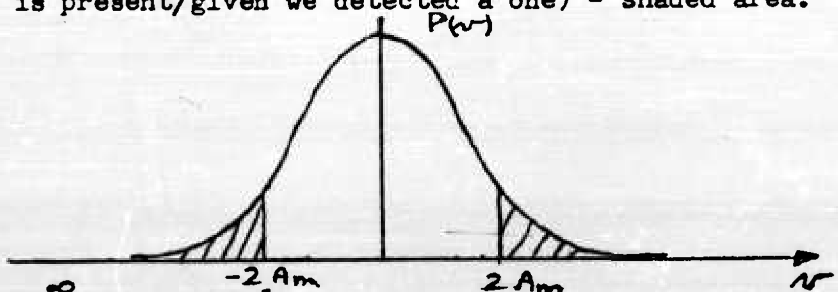


Since the amplitude of diphase one's is 12 db greater than the amplitude of dipulse one's into the detector, the peak amplitude of the triangles = 4 Am.

Again we can write: P_e = P (detecting a one) P (a zero was present/given we detected a one) + P (detecting a zero) P (a one is present/given we detected a zero).

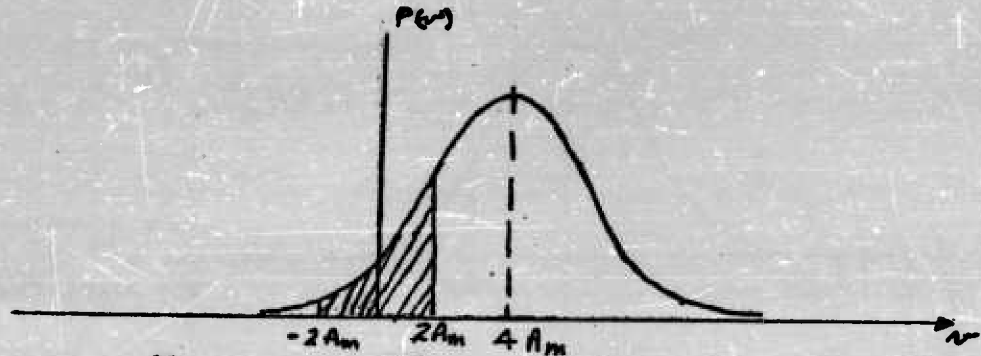
We can now obtain the conditional probabilities by examining the amplitude distribution of the noise. However, now the noise has been partially folded and added to itself due to rectification.

P (a zero is present/given we detected a one) = shaded area.



$$P_c = 2\int_{2A_m}^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma}\, e^{-\frac{v^2}{2\sigma^2}}\, dv$$

P (a one is present/given we detected a zero) = shaded area =



$$AREA \cong \int_{2A_m}^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma} e^{-\frac{v^2}{2\sigma^2}} dv$$

$$\therefore P_e = \frac{3}{2} \int_{2A_m}^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma} e^{-\frac{v^2}{2\sigma^2}} dv .$$

The effect of rectification is the 3/2 term in front of the integral. Both of the P_e's are plotted in Fig. VIII for different values of Am/σ = peak signal/rms noise.

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2. Transmission, Diphase

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A differential diphase repeater, initially developed by O. Ringlehaan, was subjected to extensive theoretical analysis. In this report dipulse and diphase systems are compared with respect to their power density spectrum, detectability, and tolerance to noise and timing extraction. Results verify that the diphase system has a higher noise threshold, less timing jitter and is a better all-around method of transmission. In addition, tests of the initial repeater models indicate that they are simpler and have lower power consumption than the present dipulse repeaters.

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1. Diphase Transmission
2. Transmission, Diphase

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III. DA Task 3B55-03-001-24

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